



RESEARCH DEPARTMENT

AN EXPERIMENTAL TRANSISTOR RECEIVER FOR V.H.F. SOUND BROADCAST RECEPTION

Report No. G-078

(1960/18)

**THE BRITISH BROADCASTING CORPORATION
ENGINEERING DIVISION**

RESEARCH DEPARTMENT

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SUMMARY

The report describes the design of an experimental transistor f.m. broadcast receiver. The circuit uses ten transistors which are currently available at moderate cost, and embodies a recently developed type of limiter and discriminator. The performance is equal, in the more important respects, to that of the better quality domestic receiver using valves.

1. INTRODUCTION

In designing a transistor broadcast receiver, there may be a tendency to make undue concessions to the small size and low power consumption of transistors to the detriment of the performance. In applying transistor techniques to the design of a v.h.f. receiver for f.m. broadcasts, it was considered desirable to avoid this tendency and to produce a receiver which, without resorting to excessive expense or complexity, had a performance equal to that of the better quality domestic valve receiver in the more important respects. In particular, adequate suppression of amplitude modulation, adequate sensitivity and reasonable fidelity of sound reproduction were regarded as essential requirements.

2. GENERAL DESIGN

The receiver is continuously tunable from 87 Mc/s to 100 Mc/s. In the initial design, consideration of the availability and cost of transistors having a useful gain at 100 Mc/s led to a decision not to use a r.f. amplifier. However, the R.C.A. type 2N247 drift-transistor was readily available; it has sufficient gain for use as an oscillator at frequencies from 77 Mc/s to 90 Mc/s and to drive a diode mixer adequately for efficient conversion of the signal to the standard i.f. of 10.7 Mc/s. Radiation of oscillator power from the aerial is minimized by using a balanced mixer circuit. The effect of supply voltage changes on the oscillator frequency is almost completely eliminated by shunting the oscillator supply by a Zener stabilizing diode. The remaining frequency drift is then quite small and is mainly caused by changes in room temperature, since the "warming-up" effects due to internal heat production are negligible.

The i.f. amplifier must have a maximum gain of about 90 dB to make up for the conversion loss of the mixer, and to supply sufficient power to the limiter; the latter can then operate satisfactorily at signal input levels down to about $30\mu\text{V}$. A stage gain of 22 dB could be achieved with adequate stability using the Mullard

type OC170 transistor and the required overall gain and bandwidth was finally obtained by cascading four such stages, using alternate single- and double-tuned coupling circuits. As miniaturization was not proposed, the i.f. amplifier could extend over some 10 in. (25 cm) of chassis to avoid unwanted feedback.

It has been emphasized elsewhere that efficient suppression of amplitude modulation is essential to reduce the effects of co-channel interference¹ and of multipath propagation.^{2,3} A simple valve receiver has been described⁴ which achieves a good performance in this respect by incorporating a discriminator combined with a "dynamic diode" limiter.⁵ The limiter operates with a self-generated bias proportional to the mean level of the applied signal and suppresses rapid changes in level. The bias voltage is also used for a.g.c. so that, for slower changes in level, the variation of a.f. output is kept small but is not completely suppressed. This enables strong and weak transmissions to be distinguished from one another when tuning. Since this combined limiter and discriminator is a passive circuit, it was adopted with little change in the transistor receiver.

A maximum a.f. output power of 1 watt, delivered to an 8 in. (20 cm) high-flux-density loudspeaker, was considered sufficient for domestic reception. A transformerless Class-AB push-pull output stage, using G.E.C. type GET115 transistors operating from a 12-volt supply, gives the requisite output into the standard loudspeaker impedance of 15 ohms. The phase-shift in these transistors at high frequencies is sufficiently small to allow about 20 dB of negative feedback to be applied to the input of the driver stage, with adequate stability. Three additional stages, one incorporating the gain control, provide the necessary gain between the discriminator output and the output stage.

For acceptable fidelity of sound reproduction, a "table-model" size of cabinet was chosen. The cabinet measures 22 in. x 16 in. x 10 in. (56 cm x 41 cm x 25 cm) and is shown in Fig. 1. As the mean power consumption of the receiver was less than 1 watt, there was no need for internal ventilation. The cabinet was, therefore, constructed as a vented enclosure with a rigid back and with internal acoustic damping. Fixed bass- and treble-lift circuits were subsequently added to the a.f. amplifier to preserve an approximately uniform axial response for the complete receiver.

A general view of the receiver chassis is shown in Fig. 2. The i.f. amplifier is built along the centre of a brass channel, 20 in. x 3 in. x 1 in. (51 cm x 7.5 cm x 2.5 cm); the r.f. and a.f. circuits are carried as separate units on each end. The left-hand knob is the on-off/a.f. gain control; the right-hand knob controls the tuning and is coupled to the scale pointer by a cord drive using flying pulleys to obtain an expanded tuning scale.

3. DESCRIPTION OF CIRCUITS

The circuit diagram is shown in Fig. 3.

3.1. R.F., Oscillator and Mixer Circuits

The r.f. and oscillator circuits are each tuned by one section of a two-gang capacitor. The effective capacitance range is reduced to a variation from

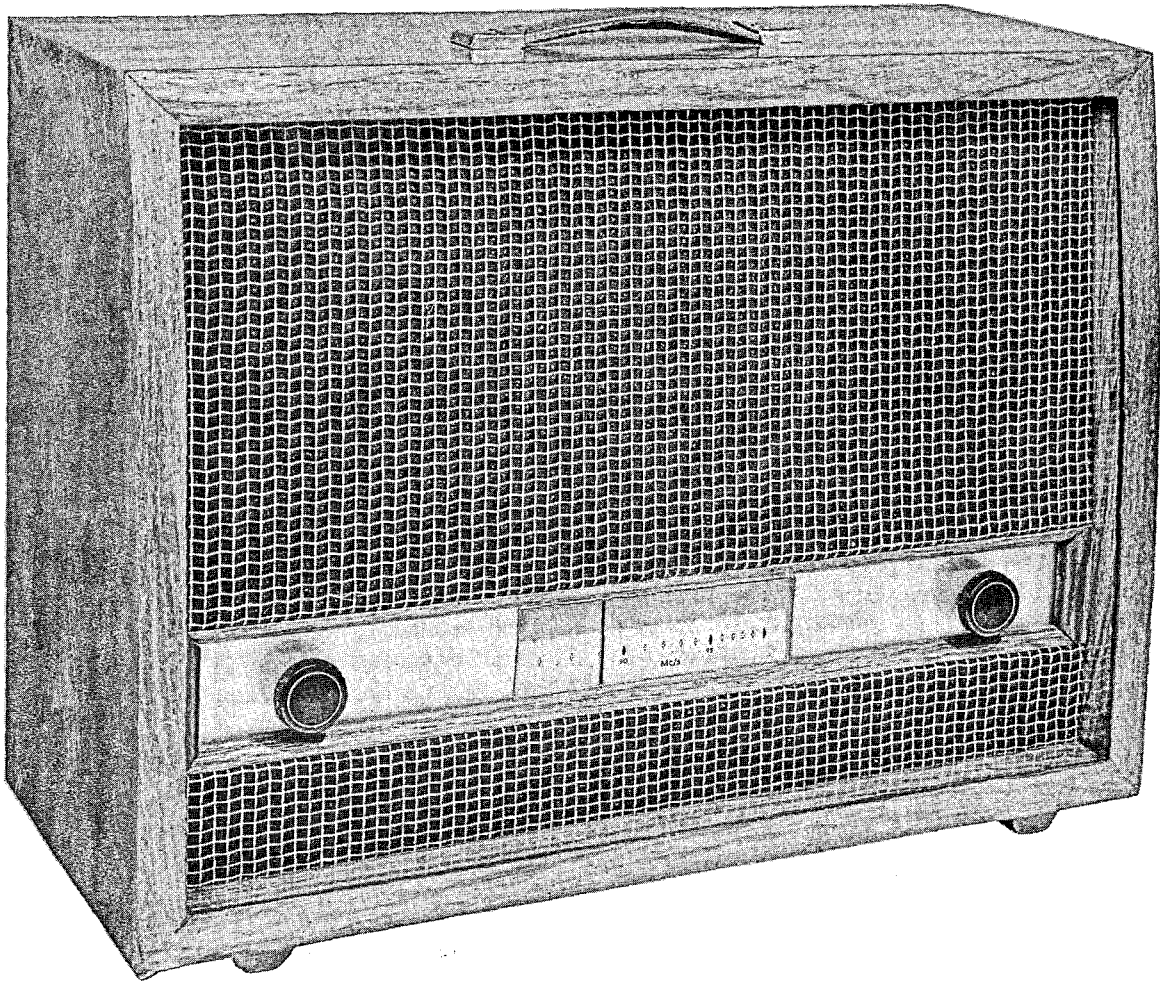


Fig. 1 - General view of complete receiver

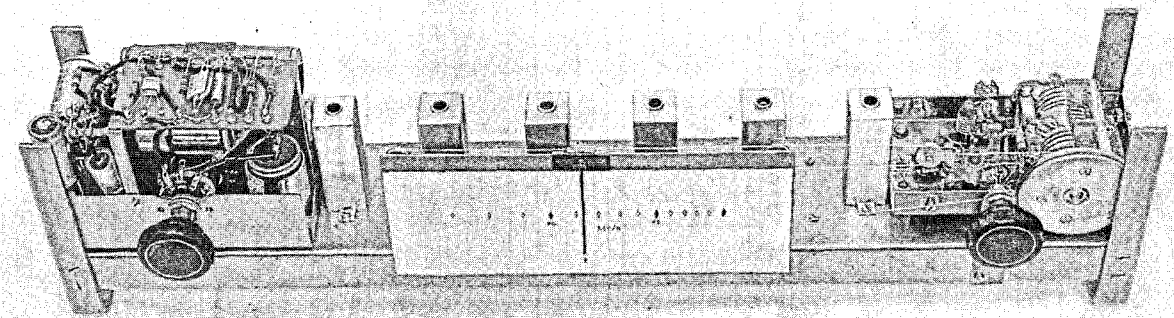


Fig. 2 - General view of receiver chassis

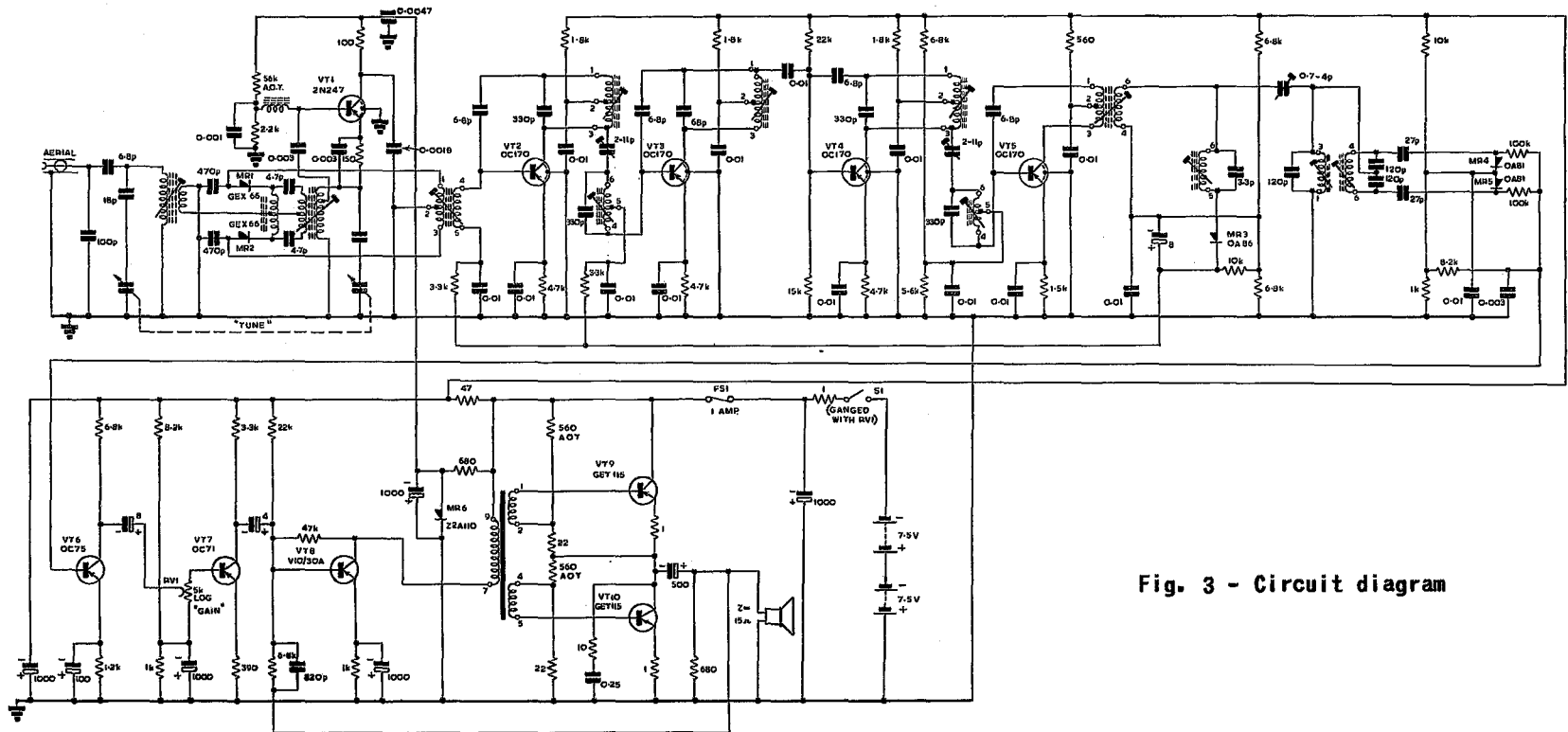


Fig. 3 - Circuit diagram

5 pF to 7 pF by adding series 18 pF capacitors and restricting the angular swing of the tuning capacitor to 150° . To assist in the tracking of the two circuits, the r.f. coil is of slightly lower inductance ($0.5 \mu\text{H}$) than the oscillator coil and is loaded by an additional capacitive network which transforms the serial impedance of 75 ohms to about 20 kilohms. The load imposed by the mixer added to the coil loss is of the same order, resulting in an effective Q-value of about 50 for the r.f. tuned circuit.

The oscillator coil is connected between the emitter and collector of a 2N247 transistor, its base being tapped in close to the emitter end. This gives the required positive feedback, bearing in mind that the phase angle between collector and emitter currents approaches 180° at 90 Mc/s. As the oscillator frequency is dependent on the collector capacitance, part of which depends on the collector voltage, the supply voltage is stabilized at 10.5V by a Zener diode, type Z2A110. The collector current is stabilized at 1 mA by an emitter resistor of 150 ohms, the base being fed via a small choke from a potential divider across the battery supply.

The mixer consists of two GEX66 point-contact diodes driven in push-pull by the oscillator; the r.f. coupling is connected between the mid-point of the oscillator coil and earth. This balanced circuit reduces the oscillator voltage appearing at the input to less than 1 mV. Direct i.f. pick-up through the mixer is minimized by the use of inductive coupling. The total direct current in the mixer is about $200 \mu\text{A}$, and is returned through the centre point of the first i.f. transformer. The output impedance of the mixer is about 600 ohms and the conversion loss of the unit at 90 Mc/s is approximately 12 dB.

3.2. I.F. Amplifier

The balanced output from the mixer is transformed to an unbalanced impedance of 300 ohms by the first i.f. tuned circuit and applied to the base of the first i.f. stage. The amplifier consists of four OC170 transistors in common-emitter connexion. The maximum possible "unilateralized" power gain is about 120 dB of which, in the interests of stability, about 30 dB is deliberately discarded. This is achieved by choosing the L/C ratios of three interstage coupling circuits to obtain a loss of about 10 dB in each coupling, using coils with unloaded Q-values of 100. The overall response is shown in Fig. 4; it is largely determined by these three circuits, one single-tuned and two coupled pairs, and is 3 dB below the maximum at about ± 90 kc/s from the centre frequency. Non-critical neutralization is effected by 6.8 pF fixed capacitors.

The first two OC170 stages operate at a maximum collector current of 1 mA, developing a potential drop of 5V across their emitter resistors. The gain of the receiver may be reduced by up to 50 dB by applying an a.g.c. voltage to the bases of these two stages, ranging from about -5V to -0.5V. The third and fourth stages operate with fixed bias; the last stage has a collector current of 3 mA in order to deliver a maximum i.f. power of about 5 mW to the limiter and discriminator circuits.

3.3. Limiter, A.G.C. and Discriminator Circuits

The collector of the last i.f. stage is connected to the primary of a tightly-coupled step-up transformer. The secondary winding is shunted by an amplitude modulation limiter in the form of a peak detector with a long-time-constant load.



Fig. 4 - Overall response preceding limiter

As the mean load resistance imposed by the limiter is approximately one-seventh of the remaining shunt resistance across the secondary winding, the limiter will function with modulation depths of up to 87%; at greater depths, the limiter will not operate at troughs of modulation. The limiting efficiency is improved by the addition of a third-harmonic filter, in series with the diode; the action of this circuit is described more fully elsewhere.^{4,6}

An a.g.c. voltage is applied to the bases of the first two i.f. transistors from one end of the limiter load resistor. The other end of the load is returned to a point on a potential divider across the d.c. supply. The a.g.c. line voltage is thus maintained at -5V in the absence of a signal, decreasing to -0.5V for large signal inputs. The a.g.c. is applied to the i.f. with no increase in the time-constant, to ensure that amplitude-modulation such as aircraft flutter is effectively controlled when the frequency is too low for the limiter to operate.

A Foster-Seeley type of phase-discriminator is coupled to the limiter by a small series capacitor which, with some of the primary reactance of the discriminator transformer, forms an impedance-inverting network.⁶ The discriminator is thus fed from a high-impedance source (as is essential for its correct operation) derived from the very low impedance presented by the limiter.

The discriminator primary voltage is applied to a capacitive centre-tap

on the secondary; the inductive coupling between primary and secondary is 1.5 times the critical value. The combined primary and secondary voltages are applied to two shunt diodes with load resistors returned through a relatively small common resistor. This resistor forms the source impedance for the first a.f. stage, carries the base bias for the transistor and, with the input impedance of the transistor and a shunt capacitance, forms the de-emphasis network. By connecting the diodes and their load resistors so that the difference-current flows in a small common resistor, the inverse has been provided of the conventional Foster-Seeley circuit which applies the difference-voltage to a large shunt resistor. The lower output impedance of the former circuit is clearly more suited to supplying a transistor amplifier. The response of the discriminator for constant voltage output from the limiter is shown in Fig. 5.

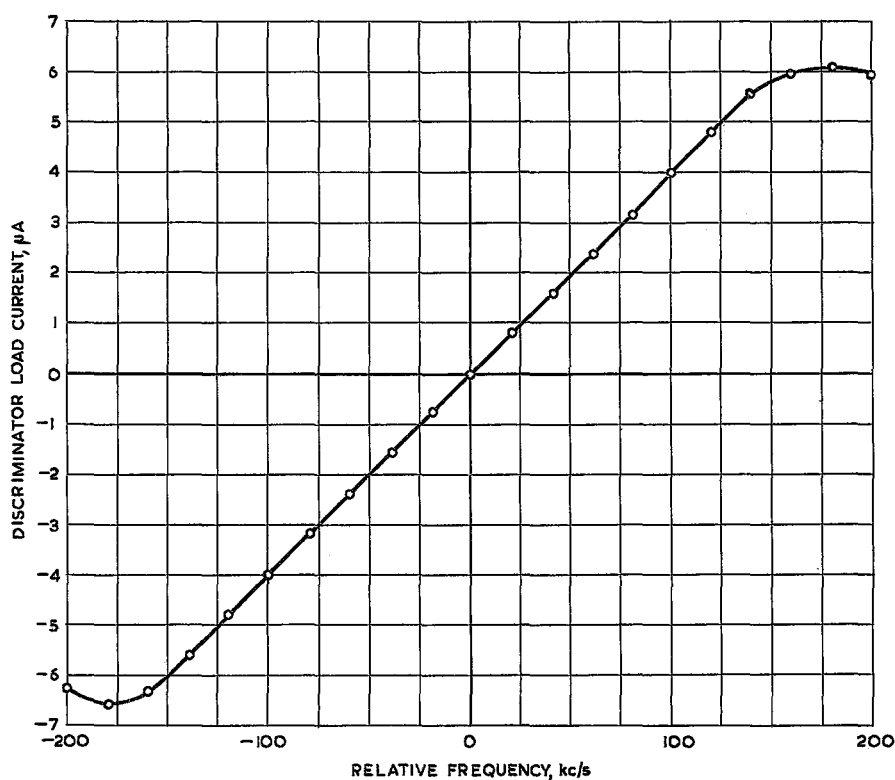


Fig. 5 - Discriminator response for constant limiter voltage

3.4. A.F. Amplifier and Output Stage

The output of the first a.f. stage following the discriminator is passed to the a.f. unit above the chassis, in which an output power of 1 watt is obtained from a minimum signal of 0.5V across the 5 kilohm gain control. There are two intermediate common-emitter stages, a coupling transformer and a pair of output transistors in economical Class-AB operation. There is no output transformer as the 15-ohm loudspeaker forms an optimum load connected (via a capacitor) between the junction of the output transistors and the chassis. The quiescent current in the output transistors is about 7 mA and some adjustment of the bias is required

4.2.2. Maximum Output Power for 10% Total Harmonic Distortion

The measured value was 1.3 watts at a supply voltage of 13.5V.

4.2.3. Modulation-Frequency Characteristic

This is shown by the full-line curve in Fig. 7. The broken curve of Fig. 7 shows the response after including the loudspeaker compensating circuits described in Section 3.5. Both curves are corrected for a $50\mu\text{s}$ pre-emphasis time constant.

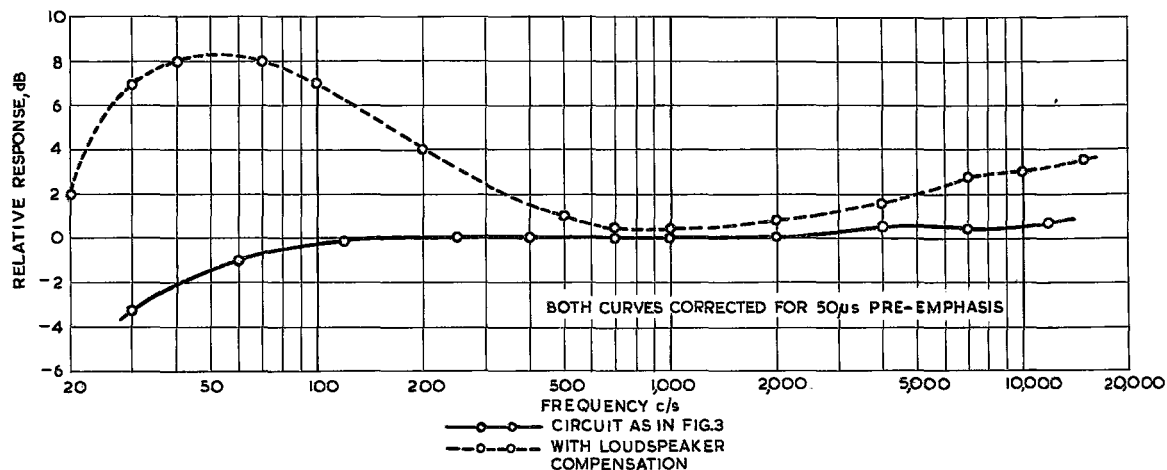


Fig. 7 - Modulation-frequency characteristic

4.3. Selectivity

The suppression ratio for an interfering signal is measured objectively as the ratio of unwanted- to wanted-signal amplitudes giving an output signal-to-interference ratio of 40 dB when the interfering signal is frequency modulated at 2000 c/s with a deviation of ± 35 kc/s.

The results for adjacent-, second- and third-channel interference (i.e. with 200, 400 and 600 kc/s frequency separations, respectively) are given in Table 1, together with the measured ratio for the image channel. The wanted carrier level in each case was 1 mV.

TABLE 1

Frequency of unwanted carrier relative to wanted carrier	-21.4 Mc/s	-600 kc/s	-400 kc/s	-200 kc/s	+200 kc/s	+400 kc/s	+600 kc/s
Ratio of unwanted- to wanted-carrier levels (dB)	+23	>+40	+40	+6	+6	>+40	+38

The i.f. suppression ratio was determined by measuring the attenuation of a signal at 10.7 Mc/s relative to that of a signal at the tuned frequency; the ratio was 68 dB when the 10.7 Mc/s input was applied via the input socket, and 60 dB when the source was connected between the outer of the socket and an external earth.

4.4. Local-Oscillator Performance

4.4.1. Local-Oscillator Drift

The frequency variation of the local oscillator is shown in Fig. 8. Since this drift is caused by a small local temperature rise in the oscillator circuit, a further measurement was made to find the result of a change of ambient temperature. This was about -14 kc/s per degree Centigrade. The drift of the discriminator centre-frequency was small compared with this.

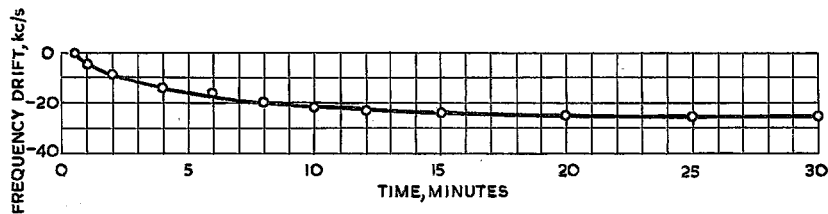


Fig. 8 - Oscillator frequency drift

4.4.2. Dependence of Local-Oscillator Frequency on Supply Voltage

The local oscillator frequency changed by +15 kc/s per volt for a supply voltage varying from 11V to 15V.

4.4.3. Local-Oscillator Radiation

In this test the voltage at the input terminals of the receiver due to the local oscillator was measured, the input terminals being terminated in 75 ohms.

The measured voltage was 0.5 mV.

4.5. Co-Channel Suppression Ratio

As for test 4.3, but with the interfering signal frequency differing from the wanted signal by less than 1 kc/s.

The measured value was -7 dB.

4.6. Suppression of Amplitude Modulation

The a.m. suppression ratio is the ratio between the output due to a carrier which is frequency modulated ± 35 kc/s at 2000 c/s and that due to a carrier which is simultaneously amplitude modulated to a depth of 40% at 2000 c/s and frequency

modulated ± 30 kc/s at 100 c/s, the 100 c/s output being rejected by a high-pass filter. The results for various input signal levels are shown in Table 2.

TABLE 2

Input Signal Level	A.M. Suppression Ratio (dB)
10 μ V	16
30 μ V	30
100 μ V	47
300 μ V	44
1 mV	49
10 mV	49
100 mV	27

4.7. Dependence of Output on Signal Level

This is shown in Fig. 9.

4.8. Impulsive Interference Performance

Fig. 10 shows the output due to impulsive interference, relative to that due to ± 35 kc/s deviation at 2000 c/s, for various input impulse amplitudes.

The measurements were made in the presence of an input carrier of 500 μ V, first unmodulated and secondly frequency modulated with ± 30 kc/s deviation at 12 kc/s.

4.9. Subjective Measurements of Selectivity and Co-Channel Suppression Ratio

For these tests the receiver was fed with two signals, a wanted signal of 1 mV and an interfering signal of controllable amplitude which was set in turn to frequencies within 1 kc/s of, and spaced by ± 200 kc/s and ± 400 kc/s from, the wanted signal.

Both signals were frequency modulated with programme in accordance with standard B.B.C. transmitter practice, the wanted programme being speech and the interfering programme light orchestral music which gave a consistently high level of modulation. The amplitude of the interfering signal was adjusted to give the following subjective grades of interference:

- JP The interference was just perceptible in the quiet passages of the wanted programme
- P The interference was perceptible in quiet passages of the wanted programme without careful listening
- SD The interference was slightly disturbing when listening to the wanted programme
- D The interference was disturbing.

The results given in Table 3 are the averages for four observers, the receiver having been tuned to give minimum output interference with the wanted and unwanted carriers within 1 kc/s, both unmodulated.

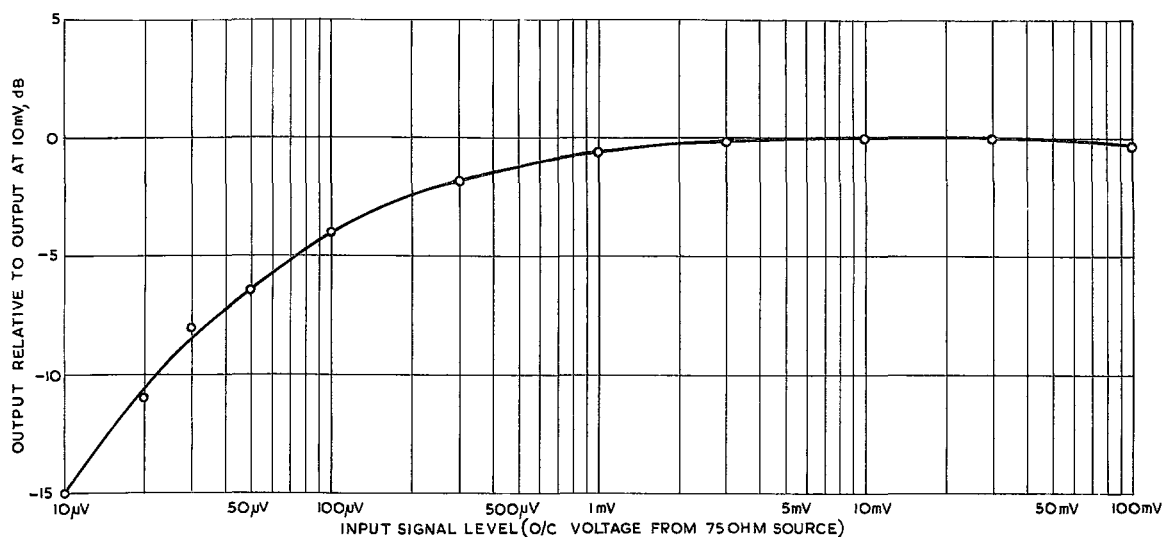


Fig. 9 - Variation of a.t. output with signal level

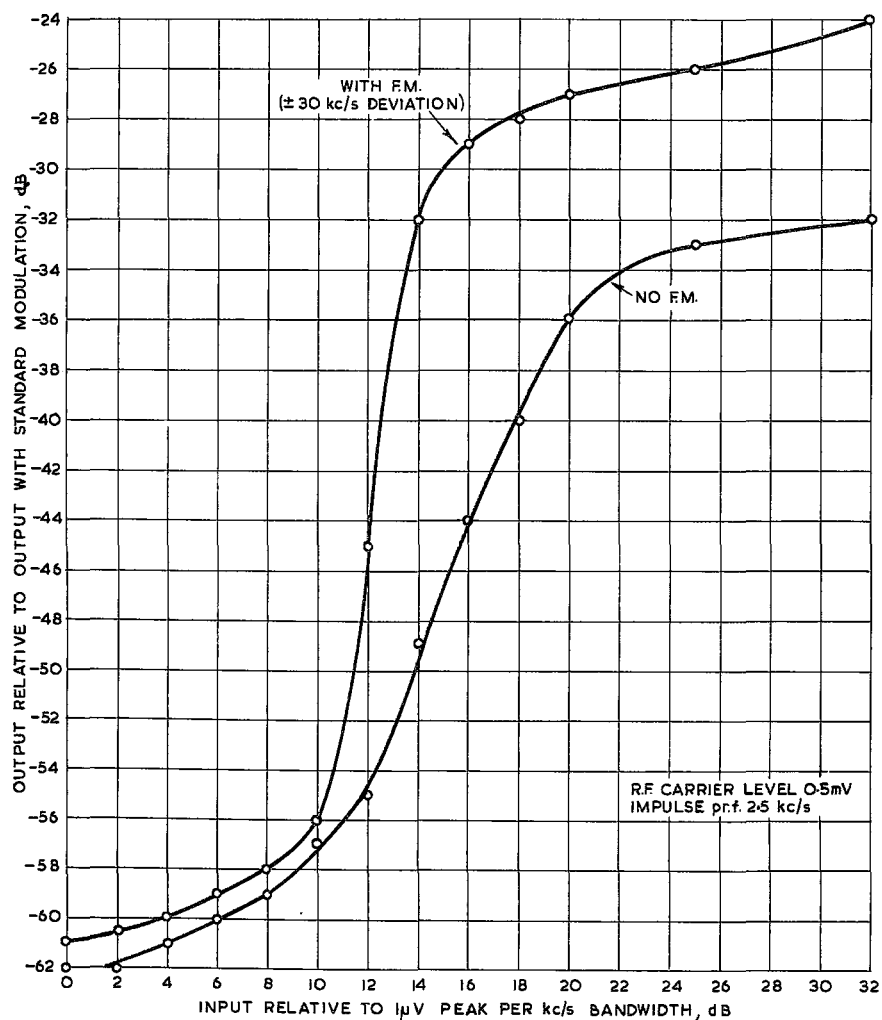


Fig. 10 - Input/output characteristic for impulsive interference

TABLE 3

Frequency of interfering signal relative to wanted signal (kc/s)		-400	-200	<+1 >-1	+200	+400
Amplitude of interfering signal (JP)		>+40	+11	-32	+8	>+40
relative to wanted signal (dB) (P)		>+40	+12	-28	+10	>+40
to give the subjective grades (SD)		>+40	+15	-24	+12	>+40
of interference (D)		>+40	+17	-18	+14	>+40

5. DISCUSSION OF RESULTS

Both from the laboratory tests given in Section 4 and from experience gained by using the receiver in more normal home conditions, its performance is entirely satisfactory in all respects save two. First, at input levels of between $10\mu\text{V}$ and $30\mu\text{V}$, the a.m. suppression ratio falls below the value of 35 dB recommended² for minimizing distortion caused by multipath propagation; this is important only when receiving unusually weak signals. Secondly, drift of the oscillator frequency becomes noticeable for large changes in room temperature. These limitations are not fundamental and methods for overcoming them are discussed below.

Although a r.f. amplifier was not used, for reasons given in Section 2, the performance of the receiver is adequate in respect of image-channel suppression and sensitivity. The standard audio output power can be obtained at an input signal level of $16\mu\text{V}$ with a signal-to-noise ratio of 43 dB; from this it can be deduced that the overall noise factor is 16 dB. There is some shortage of gain, however, in the circuits preceding the limiter, resulting in a reduced limiter efficiency at signal levels of $30\mu\text{V}$ and below, as shown in Table 2. The incomplete suppression of a.m. at these levels, in fact, allows a.m. generated in the i.f. amplifier to appear as distortion, the amount being stated in Section 4.1.2. The additional gain required to improve this situation would be better supplied by a r.f. amplifier than by the i.f. amplifier, as the gain of the latter is already as high as is convenient.

Referring again to Table 2, the a.m. suppression falls to 27 dB (somewhat less than the recommended minimum of 35 dB) at 100 mV. This is not caused by any failing of the limiter at high signal levels, but by frequency modulation of the local oscillator due to changes in the amplitude of the input signal. Although this could result in a slight increase in multipath propagation distortion and in the level of impulsive interference, these effects rarely occur at such high input levels. The addition of a stage of r.f. gain might well aggravate the above effects but, for a.m. suppression at least, an improvement would result if the a.g.c. circuit were arranged so as to reduce the r.f. gain below unity at large input levels.

The local-oscillator frequency is reasonably independent of supply voltage but does vary considerably with changes in room temperature. The rate of drift is reduced considerably by the thermal inertia of the cabinet but might be disturbing in some conditions. The effect could be compensated by the use of negative temperature-coefficient capacitors in the oscillator circuit.

Efficient rejection of impulsive and adjacent-channel interference has been greatly assisted by using a narrow i.f. pass-band, determined mainly by two

pairs of coupled circuits. These two circuits were easily incorporated in the i.f. amplifier, since four stages were required to obtain the necessary gain of 90 dB. Again, if a r.f. stage were added, it might be possible to reduce the number of i.f. stages without compromising the performance.

In spite of the high i.f. gain of the present receiver, the rejection of i.f. signals reaching the receiver by way of the aerial is satisfactory, as indicated in Section 4.3. The use of a balanced mixer with inductive coupling has ensured this high i.f. attenuation.

The apparent rise in distortion shown in Fig. 6, at low values of deviation is due principally to the presence of noise at a level 63 dB below that of the output at 40% deviation and not, as might be thought, to cross-over distortion in the output stage. This background noise is believed to arise principally from the residual f.m. noise of the local oscillator. The discriminator response, shown in Fig. 5, shows an inverse curvature at ± 100 kc/s due to a slight, unintentional over-coupling of the transformer; this will affect the rate of rise of distortion as the receiver is detuned.

Though no tuning indicator has been fitted, the receiver is relatively easy to tune on account of the a.g.c. characteristic, and the side-responses give considerably less output than the main response. In the absence of a transistor equivalent of a "magic eye" indicator, an inexpensive meter could be connected in the discriminator circuit to give a null indication of the tuning-point, if desired.

In a.f. output stages of the type used in this receiver, the loudspeaker is often connected directly between the centre-point of the output stage and the centre-point of the supply, instead of to earth via a large capacitor, as shown in Fig. 3. The former method has the advantage that the load on the supply is more uniform so that less low-frequency decoupling is necessary. Although the junction of the two batteries could have been used for this purpose, the circuit was designed so as to enable the supply to be taken from any external two-terminal source, such as an accumulator, if desired.

6. CONCLUSIONS

The development of the receiver described has confirmed the belief that, in all important respects, the performance obtained by the better quality domestic v.h.f. receiver using valves can be achieved in a straightforward design using currently-available transistors of moderate cost. By using a combined limiter and discriminator circuit⁶ of a type recently developed for a valve receiver,⁴ the a.m. suppression ratio is more than adequate. It appears that the absence of a r.f. stage has not led to any serious shortcomings, although in the present design the good a.m. suppression is not maintained below 30 μ V input. Full advantage has been taken of the high efficiency of transistors, enabling the receiver, loudspeaker and battery supply to be contained in an acoustically-treated cabinet (with no thermal ventilation), and so to provide good fidelity of sound reproduction. Changes in ambient temperature have some effect on the oscillator frequency, but these could be overcome by suitable compensation. Amplification at v.h.f. has become an economic proposition since the receiver was designed. Its use would improve the noise factor and make it easier to improve the general performance at low signal levels.

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